- 1 -

METHOD AND APPARATUS FOR REDUCTION
OF UNWANTED FEEDBACK

This invention relates to a method of and apparatus
for reduction of unwanted feedback in a system.

Unwanted feedback naturally follows from the use of
amplifiers. It has the effect of making uncontrolled
changes to the frequency response of a system. As a
result it sets a limit to the gain which can be used
before oscillation or unacceptable degradation of the
frequency response occurs.

A well-known example of such feedback is with public
address systems, where a microphone used by a person
speaking or playing a musical instrument picks up the
output of a nearby loudspeaker, giving rise to a howl
which drowns out the speech or music and renders it
unintelligible. This is sometimes termed 'howl-around'.
The system which gives rise to this is illustrated in
outline in Figure 1, which is a plan of a stage system 10
used by a speaker 12 to address an audience. His speech
is picked up by a microphone 14, and the resultant signal
amplified by an amplifier 16. A loudspeaker 18 broadcasts
the sound, some of which is picked up by the microphone
14, either directly or after reflection from the walls of
the building or stage etc. Such a sound path is indicated
by the reference 20.

It is possible to draw an equivalent circuit to this,
as shown in Figure 2. In Figure 2 is seen a signal source
30, in this case the speaker 12 of Figure 1, and an input
path 32 representing the sound path between the speaker
and the microphone 14. At the microphone the sound from

- 2 -

the speaker through input path 32 is combined in circuit
34 with sound from the unwanted feedback path 36 which
corresponds to the sound path 20 of Figure 1. The
combined signal from combiner 34 is applied to the

5   amplifier 38 in Figure 2, corresponding to the amplifier
16 and loudspeaker 18 in Figure 1. The output 39 of
amplifier 38 both constitutes the desired output and
provides the input to the feedback path 36.

A similar situation arises in a rebroadcast

10   transceiver which is designed to receive RF signals,
amplify them, and retransmit them onward on the same
frequency. While steps are taken, e.g. by using highly
directional antennas, to reduce unwanted feedback, there
is inevitably some unwanted feedback from the transmitting

15   antenna back to the receiving antenna.

Figure 3 is a generalised diagram based on Figure 2,
showing the transfer functions of the various parts of the
system. It is assumed that the input signal 30 has
spectrum I(f), the input path 32 has a frequency response

20   H(f), the unwanted feedback path 36 has a response B(f),
the amplifier 38 has a response A(f), and the output
signal 39 has spectrum O(f). The resulting overall
transfer function is:

$$\frac{O(f)}{I(f)} = \frac{A(f)\,H(f)}{1 - A(f)\,B(f)} \tag{1}$$

As the complex-valued loop gain A(f)B(f) approaches

25   (1 + j0), the system becomes unstable. If the loop
contains a dominant delay which is significant compared
with the reciprocal of the system bandwidth, then the
frequency response contains regular ripples.

To remove the effect of this feedback it needs to be

30   cancelled out. This can be done using either of the
circuits of Figure 4 or Figure 5. In Figure 4, a

compensating circuit 40 with transfer function C(f) has
its input connected to the output of the amplifier 38.
The output of the compensating circuit 40 is combined in a
combining circuit 42 with the output of the existing
combiner 34, and the output applied to the input of the
amplifier 38. Such a circuit cancels out the unwanted
feedback so long as $C(f) = -B(f)$. This is known as
neutralisation.

The overall transfer function now becomes:

$$\frac{O(f)}{I(f)} = \frac{A(f)\,H(f)}{1 - A(f)\,\{B(f) + C(f)\}} \qquad (2)$$

Commonly, some simple technique is used to ensure
that $C(f) = -B(f)$ at some spot frequency which is most
critical. Depending on the exact circuit used to ensure
that the deliberate feedback has the right amplitude and
phase this will give a more or less narrow-band solution.

Once feedback cancellation has been achieved, the
final output is what would be expected, given that the
signal has travelled the input path $H(f)$:

$$\frac{O(f)}{I(f)} = A(f)\,H(f) \qquad (3)$$

However, the circuit of Figure 4 has certain
disadvantages. For example, it might not be desired to
tap off some of the amplifier output signal into the
compensating path. Also, in the context of RF
transmission, if the processing is done at an intermediate
frequency (or at baseband), the circuit of Figure 4
requires two down-converters. This is particularly
disadvantageous if the processing in the compensating path
is digital, since each down-converter must also then be
accompanied by filtering and analogue-to-digital
conversion.

- 4 -

The circuit of Figure 5 is therefore preferred. In
this circuit the compensating path 44 with transfer
function D(f) has its input coupled in parallel with the
input of the amplifier 38, rather than being connected to
5   its output. The compensating path 44 together with the
combining circuit 42 now form a pre-corrector 46 for the
amplifier 38.

The transfer response in this case is:

$$\frac{O(f)}{I(f)} = \frac{A(f) H(f)}{1 - D(f) - A(f) B(f)} \qquad (4)$$

To cancel the unwanted feedback D(f) is chosen so
10   that D(f) = -A(f)B(f).

An example of a circuit of the type shown in Figure 5
is described in UK Patent Application GB-A-2 065 421.
This describes a rebroadcast transceiver which compensates
at baseband prior to amplification by a power amplifier.
15   The circuit described in GB-A-2 065 421 is presented in
Figure 6 of the present application. This figure is not
as such in GB-A-2 065 421, but is based on its contents.
The rebroadcast transceiver 300 is used to receive the
off-air broadcast from source 330 through channel 332,
20   combined at 334 with feedback from the transceiver output
through path 336. The main amplifier 319 of the
transceiver is preceded by a pre-corrector 346 which is
fed by a signal on an antenna line 301 from combiner 334.

The pre-corrector comprises a series circuit
25   containing a down-converter 304, a combining circuit 342,
a low-pass filter 309, an amplifier 313, a delay 314, and
an up-converter 311. A multiplicative mixer 324 is
connected to receive and multiply together the input and
the output of the delay 314. The output of the
30   multiplicative mixer 324 is applied through a low-pass
filter 326. A second multiplicative mixer 325 receives

- 5 -

and multiplies together the output of the delay 314 and the output of the low-pass filter 326. The output of the second multiplicative mixer 325 is applied to the second input of the combining circuit 324 (actually represented by a simple junction in the patent specification).

Because of the down-conversion operation, it is important to retain both phase and amplitude information in the signal. For this purpose the pre-corrector 346 operates with complex signals, but this complexity is not shown in Figure 6, for simplicity of explanation.

The circuits 324, 325 and 326 are together referred to as a "correlator" 320, the purpose of which is to compensate for feedback signals through path 336. The "correlator" 320 is stated to "correlate" unwanted frequencies sensed after filtering by the low-pass filter 309 with the unwanted frequencies present at the output of the down-converter 304 so as to effect cancellation. In fact, however, the "correlator" generates only a single output coefficient. The delay 314 is stated to be necessary in order to distinguish between the wanted signals and the feedback signal, so that the "correlator" operates only to cancel the unwanted signal. However, other known correlators may be used to produce a similar result.

This circuit can however only provide proper compensation at a single frequency, and thus is an extreme example of a narrow-band solution to the problem. It may be adequate for, e.g., military communications, but it can not provide a solution which is acceptable for broadcast quality signals. A wide-band solution is therefore necessary.

The present invention is defined in the independent claims below, to which reference should now be made.

- 6 -

Advantageous features are set forth in the dependent
claims.

Preferred embodiments of the invention are described
in more detail below. In the preferred embodiments the
5 feedback caused between the output and the input of an
amplification path is reduced by providing a delay in the
amplification path, passing through the amplification path
a signal having an auto-correlation function which is
substantially a delta function, correlating the said
10 signal before being delayed in the delay with the signal
after being delayed in the delay to produce a plurality of
correlation coefficients, modifying the signal in the
amplification path to provide a modified signal, the
modification being provided by a transversal filter
15 controlled by the said plurality of correlation
coefficients, and combining the modified signal with the
signal in the amplification path so as to reduce the
effect of the feedback. The signal having an
auto-correlation function which is substantially a delta
20 function may be an added noise signal or may be
constituted by the signal being processed itself. The
system can be used to reduce 'howl-around' in audio
situations or feedback in simultaneous coherent
re-broadcast transceivers.

- 7 -

The invention will be described by way of example with reference to the accompanying drawings, in which:

Figure 1 is a diagrammatic view of a situation in which unwanted feedback can develop;

Figure 2 is an equivalent circuit diagram for the situation depicted in Figure 1;

Figure 3 is a generalised diagram based on Figure 2 showing the transfer functions of the various parts;

Figure 4 is a diagram similar to Figure 3 showing one way of providing compensation for the feedback;

Figure 5 is a diagram similar to Figure 3 showing another way of providing compensation for the feedback;

Figure 6 is a block diagram of a known re-broadcast transceiver;

Figure 7 is a block diagram of a first feedback reduction system embodying the invention, for use with baseband signals;

Figure 8 shows part of Figure 7 in more detail;

Figures 9 and 10 are successive re-arrangements of Figure 4 illustrating the theory of operation;

Figures 11 and 12 are successive re-arrangements of Figure 5 illustrating the theory of operation;

Figure 13 is a diagram illustrating a method of measurement of finite impulse response of a circuit with a feedback loop;

Figure 14 illustrates how the finite impulse response becomes an infinite impulse response as a result of the feedback;

Figure 15 is a block diagram of a second feedback reduction system embodying the invention, for use with RF (radio frequency) signals;

Figure 16 shows part of the system of Figure 15, similarly to Figure 8; and

A

- 8 -

Figure 17 is a re-arrangement of Figure 5
illustrating the theory of operation of the second
embodiment of Figure 15.

As discussed above, a narrow-band solution to the
problem of unwanted feedback has been proposed. However,
we have appreciated that in principle a wide-band result
should be possible. The unwanted feedback is causal. In
many cases it can be modelled to sufficient accuracy as a
Finite Impulse Response (FIR) filter. It follows that
what is needed for correction is the equal-and-opposite
FIR, which is thus also causal and inherently stable. It
is necessary to determine its response, and to adjust it
if the unwanted feedback varies with time. This applies
both to the circuit configuration of Figure 4, and to that
of Figure 5. Provided that the amplifier has a finite
impulse response, D(f) is also FIR in form, though a
little 'longer' than C(f) of Figure 4.

The configuration of the compensating path 44 feeding
round through the adder 42 has the effect of making the
FIR response D(f) in effect an Infinite Impulse Response
(IIR) pre-corrector to the amplifier 38. This makes
measurement difficult, but it is nevertheless possible, as
described below.

First Embodiment

The construction of a first feedback reduction system
embodying the invention will be described with reference
to Figures 7 and 8. The theory of operation of the
circuit will then be explained subsequently.

The circuit of Figure 7 is of the type shown in
Figure 5. Thus, a signal source 30 is coupled through an
input path 32 and combined in a combiner 34 with unwanted
feedback from a feedback path 36. The output of the

combiner 34 is applied through a pre-corrector 46 to a
main amplifier 38, the output 39 of which provides the
desired output which also feeds the feedback path 36.

As noted above, such a situation pertains in many
instances where unwanted feedback arises. These include,
for example, public address systems. The circuit of
Figure 7 is suitable for operation with signals at
baseband, for example audio signals, and could be used in
a public address environment to reduce undesirable
'howl-around'.

The amplification path comprising the circuit
elements 54, 42, 56, 60, 62 and 64 contains only linear
processing elements, and thus the path provides
substantially linear processing on the signal passing
through it.

The processing in the pre-corrector 46 is essentially
done digitally, and so the input to the pre-corrector 46
is converted to digital form in an analogue-to-digital
converter 54. The output of the analogue-to-digital
converter is supplied to one input of the combiner 42, the
output of which is applied to a series circuit comprising
a variable-gain amplifier 56, the gain of which can be
controlled in response to a control signal received at a
control input 58 of the amplifier 56, a delay device 60,
and an adder 62. The output of the adder 62 is converted
back to analogue form by a digital-to-analogue
converter 64.

The pre-corrector 46 further includes a noise signal
generator 70. The output of the noise signal generator is
applied to the adder 62 in the main signal path, and is
also applied to an input X of a correlator 72. The
correlator also receives at an input Y the output of the
adder 42, that is to say the signal in the main signal

path prior to delay in the delay 60 and amplification in the variable-gain amplifier 56. The correlator 72 provides a plurality of outputs representing the correlation coefficients between the signals at its X and Y inputs. These outputs are applied to an integrator block 74. The outputs of the integrator block are then applied to control a transversal filter 76, which receives its signal input from the output of the adder 62 in the main signal path and provides its output to the second input of the adder 42. The output of the transversal filter is designed to compensate for the unwanted feedback.

The output of the correlator 72 is also applied to a controller circuit 78, described below, which controls the gain of the variable-gain amplifier 56 through its control input 58.

The construction of the correlator 72, the integrator block 74, and the transversal filter 76 are shown in more detail in Figure 8. The correlator comprises a number of stages. For simplicity of explanation, only three stages are shown, but in practice a larger number of stages will be used. The number of stages depends upon (i) the tap spacing, which is determined by the normal requirements of the Nyquist sampling thereon and the bandwidth over which the compensation is to work, i.e. at least that of the signal, and (ii) the total lengths of the response to be compensated, i.e. the extend of the convolved impulse responses of feedback path 36 and amplifier 38. Thus while only three stages are shown, it should be borne in mind that typically there will be n stages.

To the input X of the correlator 72 are connected a series of $(n - 1)$, in this case two, delays 80, each providing an incremental delay of duration T_0. The delay

time T_0 is chosen according to the desired resolution of measurement of $H(f)$ or rather its time-domain equivalent $h(t)$. The correlator 72 also includes n multipliers 82. Each multiplier 82 receives at one input a corresponding one of the X input and the $(n - 1)$ delayed versions of the X input as supplied by the delays 80, and at the other input the signal at the Y input of the correlator 72. Finally the correlator includes n low-pass filters 84, which each receive the output of a corresponding one of the multipliers 82, and provide a corresponding output of the correlator 72. The n outputs 86 from the correlator 72 represent n correlation coefficients φ_1, φ_2, . . . φ_n.

The n outputs 86 of the correlator 72 are applied to the controller 78 to control the gain of the variable-gain amplifier 56, in a manner which is described below.

The n outputs 86 of the correlator 72 are also applied through inverters 88, the purpose of which is described below, to respective integrator circuits 90 forming the integrator block 74. The integrator circuits provide an averaging function, and hold a value which is then used by the transversal filter 76. In practice, the integrators 90 and low-pass filters 84 may be combined.

The transversal filter 76 has control inputs 92 connected to the integrators 90, and a signal input 94 connected to the output of the adder 62 in the main signal path. The transversal filter 76 has n stages, corresponding to the n stages in the correlator 72, and in this simplistic example has only three stages. The signal at the input 94 is applied to the beginning of a series of $(n - 1)$ incremental delays 96 each of length T_0, similar to the delays 80. A series of n multipliers 98 each have one input coupled to one of the input 94 and the $(n - 1)$ delayed versions of it supplied by the delays 96, and the

- 12 -

other input coupled to a respective one of the inputs 92. The outputs of the multipliers 98 are then all applied to an adder 100, the output of which constitutes the output of the transversal filter 76, and is applied to the second input of the adder 42 in Figure 7. The transversal filter 76 is thus seen to be of conventional design, with a tapped delay line comprising delays 96, coefficient multipliers 98 receiving respective coefficients at inputs 92, and an adder 100 providing the output.

Theory of Operation

The circuit of Figure 7 is based on the configuration of Figure 5, which is the preferred configuration, but the principles on which it operates are also applicable to the Figure 4 arrangement. The following description of the theoretical basis for the operation of the circuit will thus cover both the Figure 4 and Figure 5 systems, and will commence with Figure 4.

It will be recalled that Figure 4 shows the equivalent circuit diagram for a system with feedback, in which a compensating path is added in parallel with the feedback path and is arranged to cancel out the feedback. The Figure 4 circuit can be arranged into the form shown in Figure 9, with certain elements added. In Figure 9, a delay 110 is added in series with the amplifier 38, and a noise source 112 is provided. An adder 114 adds the outputs of the noise source 112 and the amplifier 38, and provides the output 39 of the circuit.

The circuit of Figure 9 can then be re-arranged into the form shown in Figure 10, this again being equivalent to Figure 4 but with the same added components. (The similarity with Figure 4 is however most easily seen by considering Figure 9 first.) Since it is a simple circuit

- 13 -

re-arrangement, it need not be considered in detail. Briefly, however, it is seen that the noise signal from source 112 is injected by adder 114 into the feedback path 36 and compensating path 40, the outputs of which are

5 combined with each other and with the signal from the input path 32, and recirculated through the delay 110 and amplifier 38, to the adder 114.

Similarly, the circuit of Figure 5 can be rearranged into the form shown in Figure 11. In this instance the

10 compensating path 44 is taken before the amplifier 38, and the additional components comprise the delay 110, as with Figure 9, and also a variable-gain circuit 120 in series with the delay, a noise source 122, and an adder 124 adding the outputs of the noise source 122 and the delay

15 110 and providing an output to the amplifier 38 and to the compensating path 44. An advantage of the Figure 5 arrangement over Figure 4 is that the noise is injected at the input of the amplifier and not at the output, avoiding the need for such a high power noise source.

20 The circuit of Figure 11 can now be rearranged into the form shown in Figure 12. The noise signal from source 122 is injected by adder 124 into the composite path 130, which comprises the amplifier 38 and feedback 36, but which may be considered as one composite path as far as

25 the noise is concerned, and also the compensating path 44. The outputs of these two paths are combined with each other and with the signal from the input path 32, and recirculated through the delay 110 to the adder 124. The signal output comes from within the composite path 130,

30 but we are only interested in the noise component, and the main signal can be ignored, as seen below.

In order to cancel the unwanted feedback, it is desired, in relation to Figure 10, to make the

- 14 -

compensating path equal and opposite to the unwanted
feedback path. Thus using the terminology above, it is
desired to make $C(f) = -B(f)$. In relation to Figure 12,
it is desired to make the compensating path equal to the
5   combined effect of the amplifier and the unwanted feedback
path. Thus it is desired to make $D(f) = -A(f)B(f)$. If
this is correctly achieved, the noise signal injected at
adder 124 will be completely compensated, and the adder 42
will have zero noise contribution at its output.

10   The purpose of the feedback cancellation is to set
$C(f)$ or $D(f)$ to the required function which is necessary
to provide this cancellation. This can be achieved in
either of two ways. The first way, which may be termed
open-loop control, is to measure the distortion produced
15   in the unwanted feedback path 36 in Figure 10 or the
composite path 130 in Figure 12, and set the compensating
path accordingly. To do this it is necessary to measure
the transmission characteristic between the point X,
namely the output of the noise source 112 or 122, and the
20   point Y1, which includes the output of the feedback path
36 or the composite path 130. The point Y1 comes after
addition with the output of the input path 32 in adder 34,
because this is all that is accessible. However the input
signal is irrelevant as discussed elsewhere in this
25   specification, and can thus be ignored. The second way,
which may be termed closed-loop control, is to measure the
residual noise appearing at the output of adder 42, and
then adjust or correct the compensating path in accordance
with the amount of noise detected. Thus the error is
30   directly determined. To do this it is necessary to
measure the transmission characteristic between the same
point X, on the one hand, and the point Y2, namely the

- 15 -

output of the adder 116 [42]. Again, the signal from the input
path is irrelevant.

The transmission characteristic which is required is
the finite impulse response of the path 36 or 130,
5　respectively. However, due to the recirculation loop
through the delay 110, the response between point X and
point Y1 or Y2 will not be a finite impulse response
(FIR), but rather will be an infinite impulse response
(IIR), which is more difficult to measure.

10　Measurement of IIR

Our co-pending United Kingdom patent application No.
Published under number 5,306,012
9522150.3 entitled "Method and apparatus for determining
the response of a system containing a finite impulse
response circuit coupled with a feedback loop" provides a
15　method which can be used to measure the IIR of a circuit
of the form of Figure 10 or Figure 12. This method
involves determining the correlation between the signals
at measuring points X and Y (whether Y1 or Y2). It will
first be assumed that the input signal has zero
20　correlation with the added noise source 112 or 122. If
that is the case, the input signal and output signal can
be ignored in Figures 10 and 12 (the output signal in
Figure 12 has already been ignored in the above
discussion).

25　In accordance with the aforementioned co-pending
published under number 2,306,012
patent application No. 9522150.3, the response of a system
containing a finite impulse response (FIR) circuit coupled
with a feedback loop which introduces a loop delay T_L can
be determined. The method involves applying a test signal
30　to the system, the test signal having an auto-correlation
function which is the delta function, i.e., unity, meaning
that the signal is uncorrelated with itself. This is the

- 16 -

case with a true noise signal. In accordance with the
method the input and output of the system under test are
correlated. One of these is applied to delay means such
as to produce a series of delayed signals, but with the
total delay not exceeding the loop delay T_L. The other of
the input and output of the system under test is
multiplied with each of the delayed signals, and the
resultants of the multiplication are smoothed to provide a
degree of integration. The smoothed outputs will be
representative of the correlation coefficients of the two
inputs, and as shown in our aforementioned co-pending
patent application, will represent the FIR of the FIR
circuit.

A practical example of such a measurement system 200
is shown in Figure 13, which is based on Figure 5 of our
aforementioned patent application. The system is designed
to determine the FIR of an FIR circuit 222 with which is
associated a feedback loop having a delay 224 of delay
time T_L. The delay output is combined in an adder 226
with an input signal 212 from a signal source 230. The
aim is to determine the time-domain response $g(t)$ of the
circuit 222, when it is only possible to measure the
response of the whole circuit 220. Measurement is thus
made between the input 212 and the output 214 of the
circuit block 220. A series of individual delays 240_1,
$240_2 \ldots 240_{(n-1)}$ are provided connected in series. The
input of the first delay is connected to the input line
212 and receives the output of the signal source 230.
Associated with each delay there is a respective
multiplier 242_1, $242_2 \ldots 242_n$. Multiplier 242_1 receives
the input of the first delay 240_1 and multiplies this by
the output signal on output line 214. The output of the
multiplier 242_1 is then applied to a low-pass filter 244_1.

- 17 -

The low-pass filter effects the averaging or integrating function. The output of the low-pass filter 244_1 then constitutes the first coefficient of the correlation function φ_{xy}.

5   As shown, the signal at the input 212 of the system under test is applied to the delays 240, and the signal at the output 214 is applied directly to the other inputs of the multipliers 242. However it could possibly be the other way round, with the signal at the output 214 applied

10   to the delays 240, and the signal at the input 212 applied directly to the other inputs of the multipliers 242.

The circuit 200 of Figure 13 correlates the signals at the input and the output of the circuit block 220, and produces a correlation function which comprises a series

15   of n correlation coefficients. It is shown in our aforementioned co-pending application No.9522150.3 *Published under number 2,306, or 2* that these correlation coefficients define the FIR of the circuit 222, provided that certain conditions are met. These are that the auto-correlation function of the signal

20   from source 212 is a delta function, and that the total delay of the series of delays 240 is less than the delay in the feedback loop 224, i.e. $Nt_0 < T_L$. A degree of smoothing will also be desirable if not necessary.

The overall response of the circuit block 220 will

25   not be of finite impulse response (FIR) form, but will clearly be of infinite impulse response (IIR) form. However, the present inventor has appreciated that it is made up of successive convolutions of g(t) with itself, so that the first part of the overall response of circuit

30   block 220, over the period $0 < t < T_L$, will be identical to g(t). This assumes that g(t) is both causal and strictly limited in extent (time), such that g(t) = 0 for t < 0 and for $t \geq T_L$.

- 18 -

The manner in which the overall impulse response the is made up of the superposition of a series of 'echoes', that is to say successively more smeared-out versions of g(t), is shown in Figure 14. For the initial period 0 to T_L, there is nothing coming round the feedback loop, so the overall response is seen to be g(t). For the next period between T_L and $2T_L$, the overall response is equal to g(t) $\otimes$ g(t). For the third period between $2T_L$ and $3T_L$, the overall response contains the component g(t) $\otimes$ g(t) $\otimes$ g(t) as well. This continues for further time periods. However, it is also seen that, provided consideration is confined to the time period 0 to T_L, the form of g(t) can be determined simply from the overall response.

That is to say that with a system of the form shown in Figure 13, it is possible to measure all that is necessary in order to characterise it fully, by measuring its impulse response over a finite range, rather than for the infinite duration of its impulse response.

Figure 13 has shown the feedback path 224 as containing a simple delay. It can be shown that the principles apply equally well in the case where the feedback loop contains a delay in series with a circuit having some other response, for example having an impulse response r(t). All that is necessary is that the delay T_L exceeds the finite duration of the convolution of g(t) and r(t).

Inspection of the circuits of Figures 10 and 12 will reveal that, so far as the noise signal is concerned, they are of the form of the circuit 220 in Figure 13. Thus the correlation method of our aforementioned application can be used to determine the response of the circuits in Figures 10 and 12 which correspond to circuit 222,

- 19 -

notwithstanding that these circuits can not be isolated
from the feedback loop. This is what is achieved by the
correlator 72 in Figures 7 and 8. The correlator 72
measures the FIR of the circuits in Figures 10 and 12 as
5   follows:

	Between X and Y1	Between X and Y2
Figure 10	$B(f)$	$B(f) + C(f) = E(f)$
Figure 12	$A(f)B(f)$	$A(f)B(f) + D(f) = E(f)$

Figures 7 and 8 describe the preferred arrangement
based on Figure 12 in which measurement is taken between X
10   and Y2. Those skilled in the art will appreciate that all
the possibilities just enumerated may however be used;
the alternative point Y1 is marked on Figure 7.
Nevertheless a system in which the error is measured is
preferred for the following reason. As noted in our
15   aforementioned co-pending application, the response of the
FIR circuit can be determined from the correlation
function provided that the auto-correlation function of
the test signal is a delta function. Nevertheless, the
auto-correlation function will still introduce an unknown
20   scaling factor. Where a noise-like signal is generated,
as in Figure 7, this may not be too much of a problem, but
it is a definite problem with the second embodiment
described below.

Thus, if correlation is used to measure the unwanted
25   feedback path in isolation, such as by using the measuring
points X and Y1, the shape of the impulse response of the
unwanted feedback path will be determined, but the exact
amplitude will be unknown. It is therefore difficult to
set the corrective feedback to be equal-and-opposite. On
30   the other hand, if the error response is measured by using
the measuring points X and Y2 so as to measure the net

residual feedback after correction, then uncertainty as to its amplitude is less important. So long as the sense (sign) of the error is correctly determined, and if the shape is known, the appropriate correction can be applied.

Using this error type of measurement, what is measured is the error function, which in the time domain can be written:

$$e(t) = a(t) \otimes b(t) + d(t).$$

The aim is to make $e(t)$ as near as possible to zero. The operation proceeds in steps. The correction $d(t)$ is implemented and the error $e(t)$ is measured. This measurement is made in a sampled fashion with sampling frequency $f = 1/T_0$. The correction filter has n taps, and there are in fact n independent control loops, one for each tap. Each updates its tap weight by subtraction of an amount proportional to the corresponding element of the error measurement. For each of the n control loops, one can say:

$$d_{new} = d_{old} - \Delta$$

where $\Delta = \beta\gamma_n e$.

In this β is the unknown scaling factor in the correlation measurement, and γ_n is a factor introduced deliberately to control the performance of the control loop. The above relation will converge (assuming the feedback function stays constant) provided that $\beta\gamma_n < 1$. Thus if there is some idea of the value of β, then a sensible value of γ_n can be chosen. For rapid convergence γ_n should be as large as possible, and thus approach $1/\beta$.

The subtraction of the error element Δ shown in the above equation is illustrated in Figure 8 by the inclusion of the inverters 88. In practice the necessary inversion can be achieved at any of several points in the loop and

- 21 -

will probably be the result of circuitry included in the loop for some other purpose.

Thus, in summary, Figure 13 shows that the circuit 220 can be measured in such a way as to determine the response of the circuit 222 having an associated feedback loop 224, notwithstanding that the circuit 222 cannot be isolated from the feedback loop for measurement purposes. Figures 10 and 12 show that the circuits of Figures 4 and 5 can be regarded as containing circuits of the form of the circuit 220. In the case of Figure 10, one can regard the feedback path 36 and the delay 110 as constituting the circuit 220 (open loop control), or one can regard the parallel combination of the feedback path 36 and the compensating path 40 with the delay 110 as constituting the circuit 220 (closed loop control). In the case of Figure 12, one can regard the composite path 130, which contains the feedback path 36, and the delay 110 as constituting the circuit 220 (closed loop control), or one can regard the parallel combination of the composite path 130 and the compensating path 44, with the delay 110, as constituting the circuit 220 (open loop control).

In Figure 7, the added delay 60 corresponds to the delay 110. Having measured the response of the circuit 36 of Figure 10 or the circuit 130 of Figure 12 by the method just described, the coefficients thus obtained are used to adjust the characteristic of the compensating path 40 in Figure 10 or the compensating path 44 in Figure 12. This is achieved by means of the transversal filter 76 in Figure 7 which provides the compensating path 44 of Figure 12.

Thus, by the addition of the delay 60 in the main signal path it becomes possible to measure the characteristic of the feedback path by a correlation

method.   The characteristic of the transversal filter 76,
constituting the compensating path, can then be adjusted
to cancel the effect of the feedback.

Gain Adjustment

5   The purpose of the variable-gain amplifier 56 of
Figure 7, and the manner of its adjustment, will now be
described.   The variable-gain amplifier 56 corresponds to
the variable-gain amplifier 120 of Figures 11 and 12.

          The variable-gain amplifier is desirable for two
10   reasons:

          (i)   to regulate the power output, and
          (ii)   to maintain the stability of the
                       deliberate feedback loop 44.

As regards this second point, the feedback correction can
15   never be perfect, and it certainly will not be when the
circuit is first switched on.   This problem can be reduced
by storing the parameters which are in effect at
switch-off.

          In an arrangement based on Figure 4, it is possible
20   to adjust the gain within the main amplifier 38 without
upsetting the balance between the unwanted feedback
through path 36 and the deliberate correcting feedback
through path 40.   In other words, the adjustment of the
gain of amplifier 38 does not interact with the
25   measurement and feedback compensation operations.

          This is not the case with an arrangement based on
Figure 5, such as is shown in Figure 7.   Here a balance is
desired between the composite unwanted path $A(f)B(f)$, and
the added feedback $D(f)$ which should equal $-A(f)B(f)$.   In
30   this case the gain $A(f)$ of amplifier 38 is part of what is
being measured, and so it is preferred that it remains
constant.   For this reason a separate gain control in the

- 23 -

form of variable-gain amplifier 56 is introduced in the
main signal path so as to maintain the interdependence of
measurement and feedback compensation.

5   The variable-gain amplifier 56 could be positioned
elsewhere in the loop. Clearly it could follow the delay
60, rather than be before it, but it could also be located
in other positions.

     The basic setting of the gain α of the amplifier 56
can be chosen so as to give a value for α for the whole
10   main signal path through the pre-corrector 46 as desired,
in the manner discussed above.

     One possible approach is to start with the value of α
relatively small, that is with the gain low. Correlation
measurements are then made, and they are used to reduce
15    $E(f) = A(f)B(f) + D(f)$. Then α can be increased and a new
set of measurements made. This is repeated until the
desired power output can be achieved without loss of
stability.

     The controller 78 provides this control. It monitors
20   the outputs of the correlator 72. Precisely what it
monitors and how it responds to it is not that important;
it could simply form a weighted average of the magnitude
of the correlation coefficients. It does not need to
receive all the outputs of the correlator, but could
25   operate with selected terms only.

     Finally, the amplifier 56 can form part of an
automatic gain control (AGC) system, tending to keep the
signal within the range of the converters 54 and 64, and
preventing any attempt to radiate more power than the
30   amplifier 38 can produce. Perfect cancellation is never
achievable, but the use of the amplifier 56 allows the
cancellation to be made good enough.

- 24 -

Second Embodiment

The construction of a second feedback reduction
system embodying the invention will now be described with
reference to Figures 15 and 16.

5   The circuit of Figure 15 is also of the type shown in
Figure 5. Thus, a signal source 30 is coupled through an
input path 32 and combined in a combiner 34 with unwanted
feedback from a feedback path 36. The output of the
combiner 34 is applied through a pre-corrector 46 to a
10   main amplifier 38, the output 39 of which provides the
desired output which also feeds the feedback path 36.

The system of Figure 15 may be used in simultaneous
coherent re-broadcast transceivers, also known as active
deflectors. Such transceivers may be used by broadcasting
15   authorities as relay stations to boost local reception in
difficult areas. They could also be designed for lower
power output so as to be usable in a single building, such
as a house, to allow reception in any room in the house
without the need for a roof-mounted aerial with an outlet
20   in every room.

There are two main differences between the system of
Figure 15 and that of the first embodiment of Figure 7.
The first is that the received signals are at RF
frequency, and not at baseband, and are converted to
25   baseband for the correction processing and converted back
to RF afterwards. This means that the phase as well as
the amplitude of the signals has to be maintained, leading
to the processing of complex signals, that is signals
which may be regarded as having real and imaginary parts.
30   The in-phase and quadrature channels which are necessary
are not shown in Figure 15, in order to keep the figure
simple, but will be well understood by those skilled in
the art.

- 25 -

The second main difference is that there is no added noise signal. Instead, the input signal is itself used as the test signal. This is discussed in more detail below.

The pre-corrector 46 of the feedback reduction circuit of Figure 15 receives broadcast signals at RF frequency, and a down-converter 52 first converts them to baseband. As just noted, in order that the RF signals can be correctly regenerated at the output of the pre-corrector, it is necessary for the amplitude and phase of the down-converted signals to be retained. This means that the down-converter 52 produces two output signals, namely a 0° or in-phase signal and a 90° or quadrature-phase signal. These complex signals are processed through two parallel channels, only one of which is shown. The person skilled in the art will be familiar with the processing of complex signals in the form of real and imaginary parts, and so a more detailed description is not given here.

The down-converter 52 needs to be tuned to the RF signal being received. If the system is to be used to different RF channels, then the tuning will need to be changed (unless the system is sufficiently broad-band). When used in a low-power domestic transceiver configuration, the corrector can be arranged to be re-tuned by detecting the use of an infra-red remote-control handset of a broadcast receiver being supplied by the transceiver. Alternatively, the transceiver can be coupled to a control outlet on the receiver.

The amplification path comprising the circuit elements 52, 54, 42, 56, 60, 62, 64 and 66 contains only linear processing elements, and thus the path provides

substantially linear processing on the signal passing
through it.

The processing in the pre-corrector 46 is essentially
done digitally, and so the input to the pre-corrector 46
is converted to digital form in an analogue-to-digital
converter 54. The output of the analogue-to-digital
converter is supplied to one input of the combiner 42, the
output of which is applied to a series circuit comprising
a variable-gain amplifier 56, the gain of which can be
controlled in response to a control signal received at a
control input 58 of the amplifier 56, and a delay device
60. The output of the delay device 60 is converted back
to analogue form by a digital-to-analogue converter 64,
and is then converted back to RF frequencies by an
up-converter 66.

There is no separate noise generator, and instead the
output of the delay 60 is directly applied to an input X
of a correlator 72. The correlator also receives at an
input Y the output of the adder 42, that is to say the
signal in the main signal path prior to delay in the delay
60 and amplification in the variable-gain amplifier 56.
The correlator 72 provides a plurality of outputs
representing the correlation coefficients between the
signals at its X and Y inputs. These outputs are applied
to an integrator block 74. The outputs of the integrator
block are then applied to control a transversal filter 76,
which receives its signal input in a manner described
below from the main signal path and provides its output to
the second input of the adder 42. The output of the
transversal filter is designed to compensate for the
unwanted feedback.

The output of the correlator 72 is also applied to a
controller circuit 78, similarly to Figure 7, which

controls the gain of the variable-gain amplifier 56
through its control input 58.

The construction of the correlator 72, the integrator
block 74, and the transversal filter 76 can be as shown
5   in Figure 8, subject to the requirement for processing of
complex signals. However, some simplification can be
achieved as is shown in Figure 16, which recognises that
the signal input to the transversal filter 76 is now the
same as the signal to input X of the correlator 72.
10   Therefore only one series of delays T_0 is employed.

In Figure 16, the correlator again comprises a number
of stages. For simplicity of explanation, only three
stages are shown, but in practice a larger number of
stages will be used. The number of stages depends upon
15   (i) the tap spacing, which is determined by the normal
requirements of the Nyquist sampling thereon and the
bandwidth over which the compensation is to work, i.e. at
least that of the signal, and (ii) the total lengths of
the response to be compensated, i.e. the extend of the
20   convolved impulse responses of feedback path 36 and
amplifier 38. Thus while only three stages are shown, it
should be borne in mind that typically there will be n
stages.

To the input X of the correlator 72 are connected a
25   series of (n - 1), in this case two, delays 80, each
providing an incremental delay of duration T_0. The delay
time T_0 is chosen according to the desired resolution of
measurement of H(f) or rather its time-domain equivalent
h(t). The correlator 72 also includes n multipliers 82.
30   In this case where the down-converter 52 is used, so that
complex signals have to be handled, the multipliers 82 are
complex multipliers and are therefore designated with an
asterisk in Figure 16. Each multiplier receives at one

input a corresponding one of the X input and the (n - 1)
delayed versions of the X input as supplied by the delays
80, and at the other input the signal at the Y input of
the correlator 72. Finally the correlator includes n
low-pass filters 84, which each receive the output of a
corresponding one of the multipliers 82, and provide a
corresponding output of the correlator 72. The n outputs
86 from the correlator 72 represent n correlation
coefficients φ_1, φ_2, . . . φ_n.

The n outputs 86 of the correlator 72 are applied to
the controller 78 to control the gain of the variable-gain
amplifier 56, in a manner similar to that described above
in relation to Figure 7.

The n outputs 86 of the correlator 72 are also
applied through the inverters 88 to respective integrator
circuits 90 forming the integrator block 74. The
integrator circuits provide an averaging function, and
hold a value which is then used by the transversal filter
76.

The transversal filter 76 has control inputs 92
connected to the integrators 90. The transversal filter
76 has n stages, corresponding to the n stages in the
correlator 72, and in this simplistic example has only
three stages. A series of n multipliers 98 each have one
input coupled to one of the input X and the (n - 1)
delayed versions of it supplied by the delays 80, and the
other input coupled to a respective one of the inputs 92.
Thus the delays 80 form part of the transversal filter as
well as the correlator. The outputs of the multipliers 98
are then all applied to an adder 100, the output of which
constitutes the output of the transversal filter 76, and
is applied to the second input of the adder 42 in Figure
15. The transversal filter 76 is thus seen to be of

conventional design, with a tapped delay line comprising delays 80, coefficient multipliers 98 receiving respective coefficients at inputs 92, and an adder 100 providing the output; notwithstanding that the delays 80 are used in common with the correlator 72.

As noted above, whether the arrangement of Figure 8 is used or that of Figure 16, the multipliers 82 are complex multipliers. What in fact is required, from a theoretical standpoint, is that when the multiplier multiplies its two inputs X and Y, one of them should in fact be the complex conjugate. That is to say the multiplier should form X*Y, where the asterisk indicates the complex conjugate. Two processes thus seem to be necessary to produce X*Y, namely (i) form the complex conjugate X* from X, and (ii) perform the complex multiplication X*Y.

However it can be shown that simply changing the sign of certain terms calculated in a complex multiplication achieves the same result as if one of the inputs had first been conjugated, while requiring no additional hardware. That is:

If $X = a + jb$

and $Y = c + jd$

where a, b, c, and d are all real, and j is the square root of minus one, then the product:

$XY = (a + jb)(c + jd) = (ac - bd) + j(ad + bc)$.

The corresponding product, in which the complex conjugate of X is taken, is however:

$X^*Y = (a - jb)(c + jd) = (ac + bd) + j(ad - bc)$,

that is, the same as XY but with the signs of two terms both changed. To form either XY or X*Y directly from complex inputs X and Y thus requires the same number of

operations. These are four real multiplications, one
addition and one subtraction.

In Figure 16 the asterisk (*) by one of the inputs of
the multipliers 82 denotes an input which is, in effect,
conjugated before the complex multiplication process,
while requiring no more complexity than a normal complex
multiplier. In principle, XY* could be used instead of
X*Y.

Theory of the second embodiment

In certain circumstances the input signal may itself
be sufficiently noise-like to fulfil the requirement of
having an auto-correlation function that is substantially
equal to the delta function. In that case, the
correlation measurement can be achieved without the need
to add an extra test signal.

Figure 17 shows how the circuit of Figure 5 can be
re-arranged with the addition of the variable-gain-circuit
120 and loop delay 110, in similar manner to Figures 11
and 12, but without the addition of the noise source. It
is seen that a circuit of the form of the circuit 220 of
Figure 13 still results. One can regard the composite
path 130 and the delay 110 as constituting the circuit 220
(open loop control), or one can regard the parallel
combination of the composite path 130 and the compensating
path 44, with the delay 110, as constituting the circuit
220 (closed loop control).

In Figure 15, the added delay 60 corresponds to the
delay 110. Having measured the response of the circuit
130 of Figure 17 by the above method, the coefficients
thus obtained are used to adjust the characteristic of the
compensating path 44. This is achieved by means of the

- 31 -

transversal filter 76 in Figure 15 which provides the
compensating path 44 of Figure 17.

In this instance the X input to the correlator 72 is
not, of course, the output of the noise source 70 as in
Figure 7, but is the output of the delay 60. The Y input
is taken from Y1 or Y2 depending on whether open-loop or
closed-loop control is used; Figure 15 shows closed-loop
control, which is preferred.

Again it is seen that by the addition of the delay 60
in the main signal path it becomes possible to measure the
characteristic of the feedback path by a correlation
method. The characteristic of the transversal filter 76,
constituting the compensating path, can then be adjusted
to cancel the effect of the feedback.

Using the wanted signal as the test signal itself
gives improved signal-to-noise ratio of both the
measurement, and the resulting output signal, as no noise
has to be added. There is also the instrumental advantage
of not needing an extra circuit in the form of the noise
generator.

Gain adjustment in the Second Embodiment

The signal X(f) appearing at point X can be seen from
Figure 17 to be:

$$X(f) = \{ I(f) H(f) + [A(f) B(f) + D(f)] X(f) \} \alpha e^{-j\omega T}$$

$$\ldots \quad (5)$$

This can be re-arranged to give:

$$X(f) = \frac{\{ I(f) H(f) \} \alpha e^{-j\omega T}}{[1 - \alpha e^{-j\omega T} \{ A(f) B(f) + D(f) \}]} \quad (6)$$

In order for the auto-correlation function of the
signal to be as close as possible to a delta function,

X(f) should be as flat as possible across the frequency
spectrum. However, X(f) will not be flat if any of the
following are not flat:

      (i)   the original wanted-signal spectrum I(f);

5   (ii)   the input path H(f); or

      (iii) the denominator of expression (6).

The last of these, (iii), represents the extent to
which the spectrum is shaped by the loop in the system.
Its influence can be minimised both by keeping the gain α
small, and by improving the balance between the unwanted
and compensating feedback so that A(f)B(f) + D(f) tends to
zero.

The best method of operation is thus to follow an
iterative procedure as follows. Initially the value of α
is set small, that is the gain is very low. Correlation
measurements are then made and an initial set of
correlation coefficients are obtained. The compensating
path is adjusted to bring D(f) closer to -A(f)B(f), i.e.
A(f)B(f) + D(f) will be reduced. Then α is increased, and
the measurement step repeated. This will further improve
the compensation, and thus α can be increased again.
These steps are repeated, until the desired power output
can be achieved without loss of stability. Provided that
the initial measurements yield close enough results, the
process will converge to the correct setting.

Implementation

The circuits illustrated in Figures 7, 8 and 15, 16
have been described in terms of a hardware embodiment,
with discrete circuits. It will be appreciated that the
circuitry may however be implemented in part or wholly in
software. In this case the figures should be regarded as
equivalent to flow charts.

- 33 -

The circuits of Figures 15 and 16 may be implemented for use with orthogonal frequency division multiplex signals, as described in our ~~United Kingdom patent application No. 9522198.2 entitled "OFDM Active Deflector" and our co-pending European and United Kingdom Applications of even date claiming priority therefrom.~~

The most difficult part of the above procedure is at switch-on. It is possible to overcome this problem by designing the system so that it is switchable between two states, in the first of which there is an added noise generator as in Figure 7, and in the second of which the wanted signal is used as the test signal, as in Figure 15. Initially measurements are made with added noise using the Figure 7 configuration (it being appreciated that the down converter 52 and up converter 66 are added to Figure 7) and the gain of the variable-gain amplifier is set to zero. Setting α to zero ensures that the system cannot oscillate at switch-on, and the correction can be trimmed to balance out the unwanted feedback. Once this is achieved the noise generator can be turned off or disconnected and the system re-configured as in Figure 15. Now the wanted signal is used for measurements and the value of α slowly increased. As the loop now starts off in an in-balance condition, any errors should be small, so that the spectrum of the measurement signal should be sufficiently flat. This will ensure that the above iterative process converges on the correct setting as α is increased.

Reduction of the added delay

As is seen from the foregoing, insertion of a delay T_L within the loop permits the unambiguous measurement of either the unwanted feedback itself or the error caused by

- 34 -

the unwanted feedback when combined with the wanted signal. However, the same delay T_L is introduced into the passage of the wanted signal. This has disadvantages in all the example referred to above. For example, with the

5   first embodiment it is disadvantageous if the sound is delayed when the audience can see the speaker or performers on stage, or hear them directly as well as through the amplifier 38.

The delay 60 is under the system designer's control,

10   and will normally be chosen to exceed the maximum delay caused in the feedback path 36. We have appreciated that, where closed-loop control is used in the correlation measurement, the delay period can be adaptively reduced.

Referring back to Figure 14, it will be recalled that

15   this illustrates the overall infinite impulse of a system comprising an FIR with characteristic $g(t)$, together with an added recirculating delay T_L which is longer than the extent of $g(t)$. Such a system is represented by the circuit block 220. The overall response of the circuit

20   block 220 starts with a distinct version of $g(t)$, followed by a series of overlapping and decaying echoes, as described above.   If T_L is shortened so that the extent of $g(t)$ is greater than T_L, then the first part of the loop's response between O and T_L no longer represents

25   a complete, distinct picture of $g(t)$. The last part of $g(t)$ will merge with the first part of the first echo between T_L and $2T_L$. Nevertheless, the first part of $g(t)$, up to time T_L, remains distinct and measurable.

If this situation pertains with one of the

30   above-described embodiments, such as in Figure 7 or Figure 15, where closed-loop control is used, the taps in the transversal filter 76 which cover the time up to T_L will still converge correctly to the required correction.

Their error measurements are never polluted by the echoes.
This assumes that the unwanted feedback path is unchanging
(or, in practice, changes relatively slowly). The taps in
the transversal filter 76 which cover the time period

5   beyond T_L will be polluted by the echoes as this first
phase of the operation takes place.

However, once the taps of the transversal filter
covering the time period O to T_L have converged, then the
error e(t) over the period $0 \leq t \leq T_L$ will be zero. This

10   means that the first part, of length T_L, of the first
'echo' will also be zero. Thus the measurements of the
error e(t) for the period $T_L \leq t \leq 2T_L$ are also now
distinct, so that the corresponding taps of the corrector
can converge. This process can continue indefinitely in a

15   'wave of correction', sweeping outwards from the origin,
for a corrector of any arbitrary length exceeding T_L.

It may be preferred not to try to correct the outer
terms until the inner terms of the transversal filter have
settled down. That is, the later corrector taps can be

20   held at zero, as their error measurements are known to be
polluted during the initial correction phase, until the
coefficients for the period O to T_L have been determined.

Thus T_L can be chosen to be shorter than the maximum
delay introduced in the feedback path and the system will

25   still function. In principle, the delay could be reduced
down to a single sampling period, in which case the
coefficients will be determined one by one.

The main disadvantages is that the process takes
longer to converge. Roughly speaking one can say that if

30   the control loops are set so they each take T_C to
converge, then for a system where the extent of g(t) does
not exceed the delay T_L, the loops can all settle
independently in parallel and overall system convergence

takes T_C. If the extent of $g(t)$ does exceed the delay T_L, and is say kT_L, then convergence will take roughly kT_C, as the loops converge in a combined sequential/parallel fashion, T_L's worth at a time. While in principle the delay T_L could be reduced to a single sample delay, in this case convergence would be strictly sequential, each tap having to wait for all the previous ones to settle first. Furthermore, there could be difficulty if, as is inevitable in practice, the auto-correlation function $\varphi_{xx}(\tau)$ of the 'test signal' departs from a pure delta function.

This is the case whether the test signal is an added noise signal or the wanted signal itself. On the other hand, a simpler correlator could be employed which only evaluated the cross-correlation for a single point at a time, rather than evaluating it for all points in parallel, as in Figures 8 and 16.

Another problem is that the ability of such a system to cope with varying unwanted feedback paths is limited. Each time one of the 'early' taps was disturbed, many or even most of the subsequent taps are disturbed too. So, the use of a delay T_L which is shorter than appears to be necessary is best suited to situations where little (or at least very slow) variation in the unwanted feedback is to be expected. The more nearly the delay T_L approaches the extent of $d(t)$, the quicker the convergence and the better the ability of the system to cope with changes in the unwanted feedback.

There may be advantages in choosing different values to T_C for the control loops of the various taps, setting it to be shorter for the early ones, and longer for the later ones. Furthermore, initial lock (on switch-on or RF-channel change) might be speeded up by starting from the last-used correction, stored for the purpose.

It has been assumed that T_L is chosen to be a fixed value. It may be possible to provide an adaptive arrangement in which T_L starts at a large value, giving quick convergence but introducing maximum delay in the signal path, and is then successively reduced.

It will be appreciated that many modifications and variations can be made to the systems described within the scopes of the appended claims.